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WIDE SPECTRUM MICROWAVE PULSE MEASUREMENT

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1. INTRODUCTION

Along with the rapidly moving technology of high power microwave pulse generation, there is a growing need to develop wide band diagnostics for measuring the instantaneous amplitude and frequency content of one-shot pulses. While means are available for measuring the spectra of bands up to 1 GHz wide, octave bands above 1 GHz present greater challenges. Commercial spectrum analyzers are of no use for one-shot pulses, so the spectrum(s) must be converted to much lower frequencies where the time-waveform can be captured, either digitally or on a CRT. Also, drawing from modern electronic countermeasures technology, smart digital RF memory (DRFM) devices which invoke digital parallel processing may be of use.

The diagnostics discussed here are not intended to be exhaustive, particularly new or revolutionary. Neither have they been tested at LLNL. They are simply offered as possible tools for use by the microwave community, "for the record." In most cases, the utility of a particular technique must be assessed through experimentation and the experience of other experimentalists. This document is intended to be a supplement to reference [1].

In all of the techniques discussed in the following, we are chiefly interested in preserving coherency (instantaneous amplitude and phase) of the entire pulse waveform as much as possible. This is in contrast to simple envelope (video) detection where all coherency is lost, and only the total power in the band is measured versus time. While information about the total power in the band is useful information, it is perhaps more important to know how the power is distributed throughout the entire frequency spectrum. Spectral information is especially crucial when the radiated pulse is considered to be a threat to electronic systems through some coupling path since such systems are usually frequency selective.

In order to achieve adequate spectral resolution using envelope detectors, it is necessary to filter the entire spectrum into many slightly overlapping sub-bands or bins only several tens of MHz wide, each branch being terminated in a detector [1]. Not only are narrow bandwidths required to provide the necessary degree of spectral resolution, but also to provide the necessary sensitivity as established by the detector's noise threshold.

2. FREQUENCY DISCRIMINATOR AND MIXING TECHNIQUES

In the following methods it is also necessary to filter the wide-band input into m slightly overlapping sub-bands or bins (Fig. 1). This is because the down-conversion process preserves the bandwidth. Each bin has a bandwidth B which can be recorded by an oscilloscope or captured by a DRFM. Of course, coherence between the bands is lost. Each band can be handled using a Polar Frequency Discriminator (PFD), or by down conversion using a local oscillator and mixer.

2.1. POLAR FREQUENCY DISCRIMINATOR (PFD)

This method appears to be a quick, simple, and inexpensive way to measure the instantaneous amplitude and frequency of pulses having relatively small bandwidths [1]. As shown in Fig. 2, a PFD consists of a polar phase discriminator (PPD) preceded by a hybrid splitter and a delay line. The PPD has four detector outputs which consists of constants (dc terms) and combinations of [cos $\phi(t)$, sin $\phi(t)$]. When fed to the differential axes of the oscilloscope, the constants drop out leaving

$$V_2 - V_1 = AB \cos \phi(t) \rightarrow \text{to x-axis} \tag{1a}$$

$$V_4 - V_3 = AB \sin \phi(t) \rightarrow \text{to y-axis} \tag{1b}$$

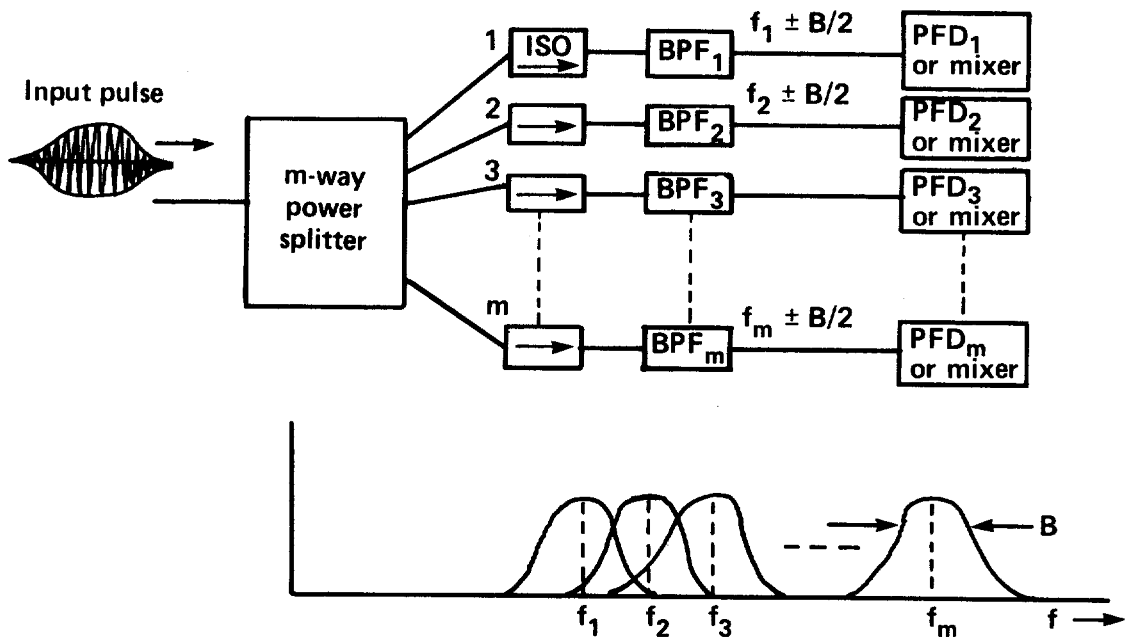

Fig. 1 The overall spectrum is filtered into m sub-bands for analysis

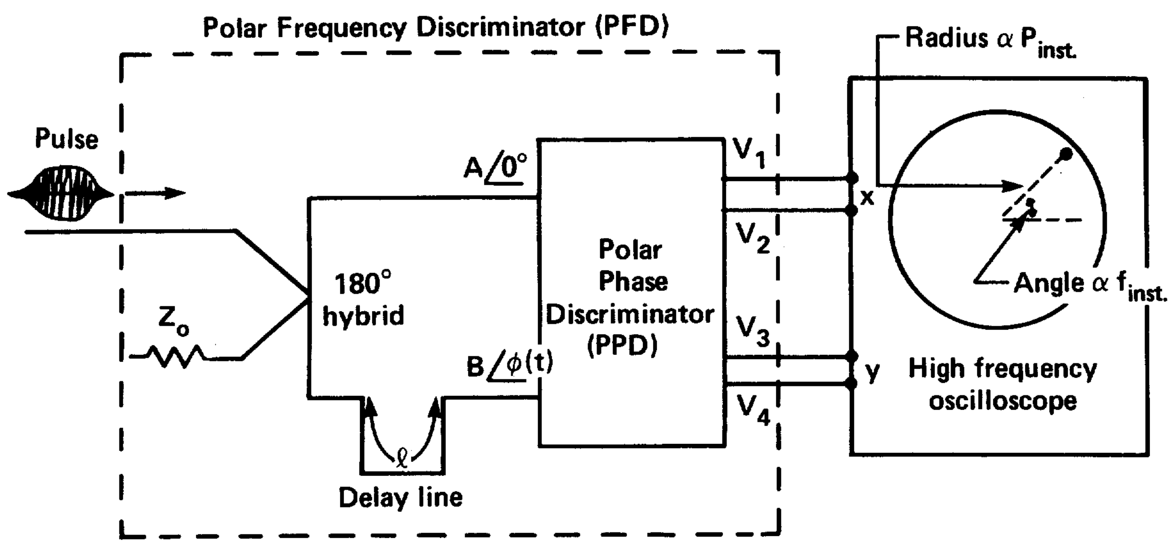

Fig. 2 Polar frequency discriminator method for measuring the instantaneous power and frequency

where $\phi(t)$ is the instantaneous phase difference between the two inputs to the PPD. Thus, using a dispersionless delay line of length ℓ to create this phase difference

$$\phi(t) = \beta\ell = \frac{2\pi f\ell}{\nu_p} = \text{const} \times f(t) \quad (2)$$

where ν_p is the assumed constant phase velocity of the delay line.

The oscilloscope will therefore display a polar plot, with the angle being proportional to instantaneous frequency while the radius (AB) is proportional to the instantaneous power. Normally, A = B except when the instantaneous pulse amplitude changes significantly in the delayed branch due to mismatches or dispersion.

PFD units are available from at least three sources (Aertech, Narda, and Anaren) in the range of 1 - 18 GHz in octave steps. In most PFD applications the displayed angle is less than 360° (e.g., 340°) over the octave band to avoid ambiguity. However, even though it is important to be able to measure the mean frequency of the pulse, it is also desirable to know how the frequency varies along the pulse. Consequently, the length of the delay line should be great enough to cause the 340° of rotation over the bandwidth (B) of the pulse rather than over the rated octave spread of the PFD. This gives better frequency resolution over the pulse. This can be done if the unit has provision for adding delay lines externally; only the Narda and Aertech units have such provision.

Calibration of the unit including the chosen delay line is accomplished using a cw swept source. The linearity of angle with frequency is typically ±7°.

The tangential sensitivity of the Narda unit is -45dBm in a 2 MHz bandwidth. At 500 MHz bandwidth, the sensitivity is reduced by 24dB, giving a dynamic range of only 6dB, assuming the detector square-law region is below -15dBm which is typical of Schottky and tunnel diodes. The PFD is usable up to +12dBm, but the response is nonlinear and calibration would be required. This would extend the dynamic range to 33dB, with a sensitivity of -21dBm.

Additional discussion and details on PFD's for single pulse diagnostics is given in [1].

The sensitivity and dynamic range are substantially improved using the homopulse or heteropulse systems as discussed in the following.

2.2. HOMOPULSE DETECTOR

Figure 3 shows a system for measuring the pulse spectrum, but it does not measure the carrier frequency (ω_0). Its advantages are:

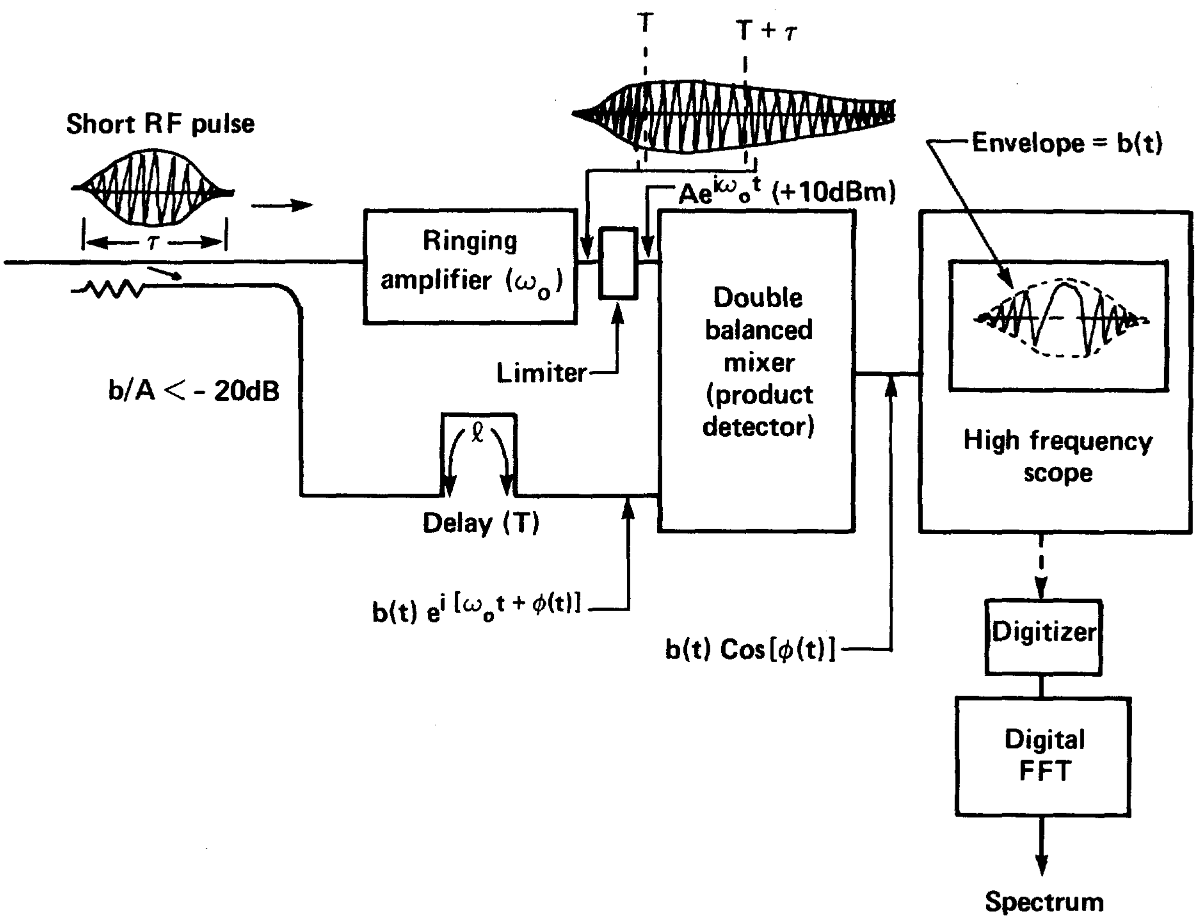

Fig. 3 Homopulse detector

(a) high accuracy over the entire pulse

(b) zero IF

(c) great sensitivity and dynamic range

Basically, the system coherently mixes a "cw" carrier RF signal (ω_0) with the delayed pulse, in much the same way as a Doppler radar operates. This carrier is generated using a high Q "ringing" amplifier which tends to sustain the oscillations for a period much longer than the pulse duration, τ. Assuming the ringing amplifier takes T seconds to stabilize, the pulse should be delayed by T to permit correlation to occur from $T<t<T+\tau$. Designing a ringing amplifier will require some literature and experimental research. Technology should exist for doing this since some high power pulsed radar transmitters tend to ring.

The ringing amplifier is followed by a limiter which provides a constant input (+10dBm) into the mixer over the correlation time. The limiter not only protects the mixer from overdriving and possible burnout, but also sets the reference power for maximum sensitivity and dynamic range. The mixer operates as a linear product (homodyne) detector [2] so its output is

$$b(t) \cos [\phi(t)] \tag{3}$$

for an instantaneous pulse of

$$b(t)\, e^{i[\omega_0 t + \phi(t)]}. \tag{4}$$

Here, $b(t)$ is the instantaneous pulse amplitude and $\phi(t)$ is the instantaneous pulse phase, relative to the reference signal $Ae^{i\omega_0 t}$ at the input to the mixer.

To minimize detector mixing errors, b/A is normally chosen to be small, say < -20dB [2].

The instantaneous frequency deviation is

$$\omega(t) = \frac{d\phi(t)}{dt} , \tag{5}$$

from which the bandwidth B can be determined. For narrow bandwidths, the detected envelope will give a true representation of b(t) only when $\phi(t)$ changes several multiples of π over the pulse length, τ. This will give multiple oscillations as shown on the oscilloscope trace in Fig. 3.

Note that the mixer's time-varying output is due to the amplitude and phase modulation of the pulse. In contrast to heterodyne mixing using a local oscillator which gives a mixer output at the IF, here the IF is zero since the "local oscillator" is actually the center carrier frequency of the pulse. The advantage of homodyne (product) detection is that the pulse spectrum is shifted down all the way to zero IF and is therefore in the range which is most easily captured. Digitizing the complex pulse waveform and taking its Fourier transform gives the voltage spectrum. Note that this is not the power spectrum, since the mixer is linear.

The sensitivity for a 500 MHz bandwidth is about -60dBm, and the dynamic range is 60dB.

If the spectrum in any one branch is reasonably symmetrical, the ringing amplifier can be made to ring at the center frequency which is presumably strongest. Then, on conversion to zero IF, the portion of the spectrum which lies in the negative frequency range would be caused to flip over to make the measured amplitude twice as large over a bandwidth of only B/2, as shown in Fig. 4a. Obviously, there are problems if the spectrum is not symmetrical since the negative frequency part that flips over will differ from that for the positive frequency part. Only the total for $f > 0$ will be observed, so we will be unable to ascertain the true shape of the whole band (Fig. 4b). Similar problems occur if the amplifier does not ring at the center frequency (Fig. 4c). In this case, the IF is not zero and the negative frequency part will flip over to give an apparent spectrum as shown by the solid line, rather than the true spectrum shown by dashed line.

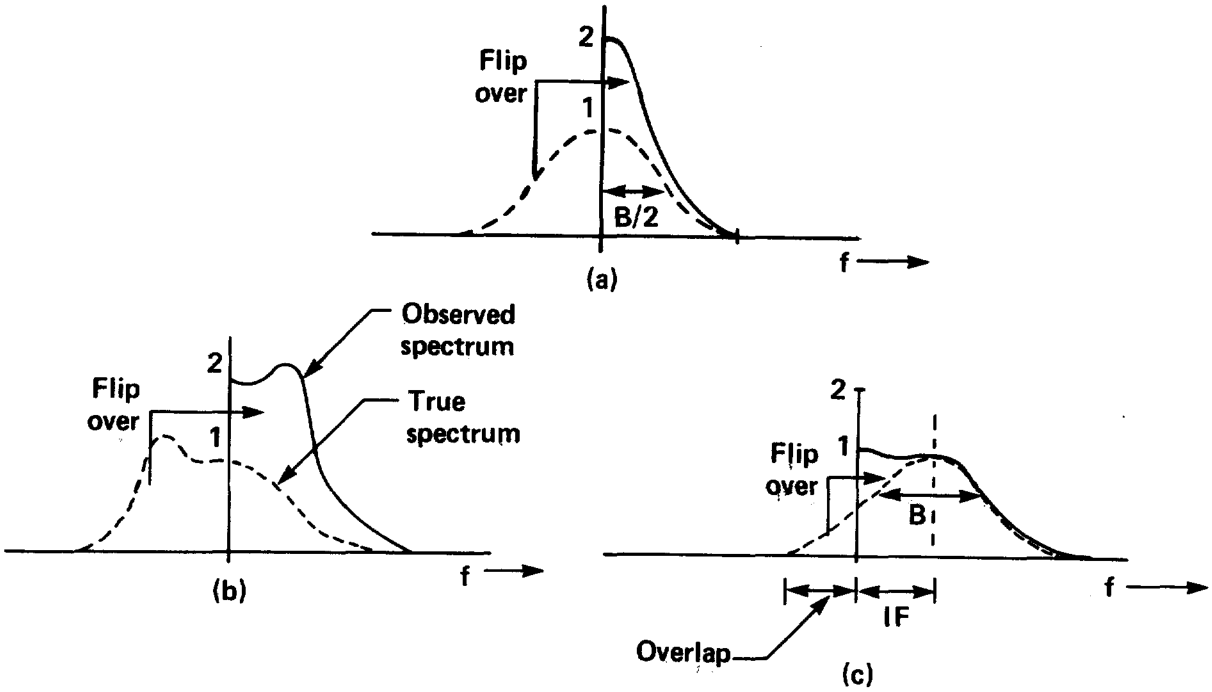

Fig. 4 Homopulse spectra
(a) symmetrical, zero IF
(b) asymmetrical
(c) symmetrical, nonzero IF

2.3. HETERODYNE PULSE DETECTOR

Instead of using a ringing amplifier to generate a "LO" at the center frequency of the channel pulse, we can use heterodyne mixing where the LO frequency is set near the lower edge of the band as shown in Fig. 5. The transformed spectrum would then be that shown in Fig. 6. The advantages of this system are:

(a) high accuracy over the entire pulse

(b) great sensitivity and dynamic range

The sensitivity of a heterodyne system having a 500 MHz bandwidth is of the order of -58dBm. Conversion is linear up to about -15dBm, giving a dynamic range of 43dB.

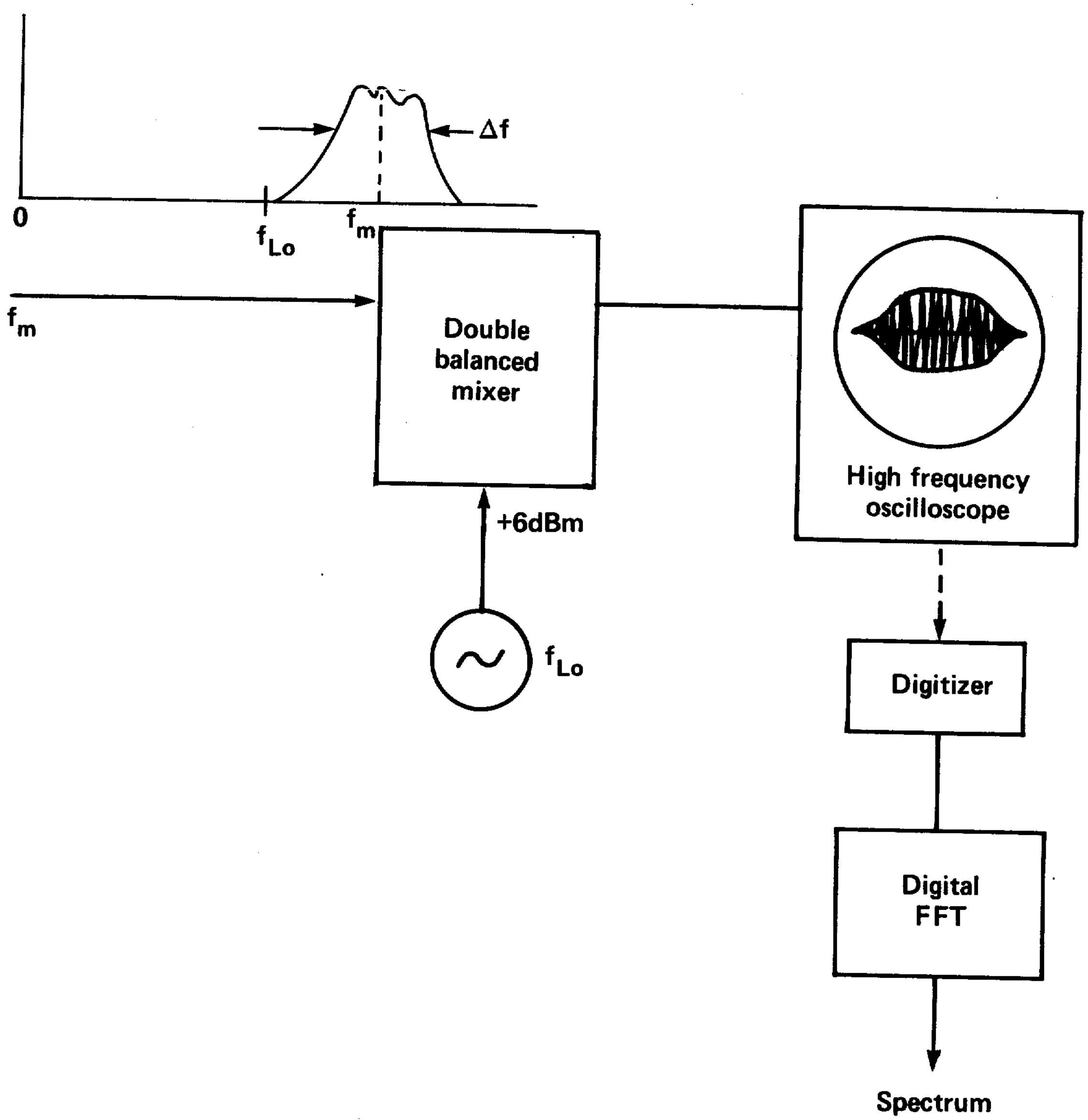

Fig. 5 Heteropulse detector

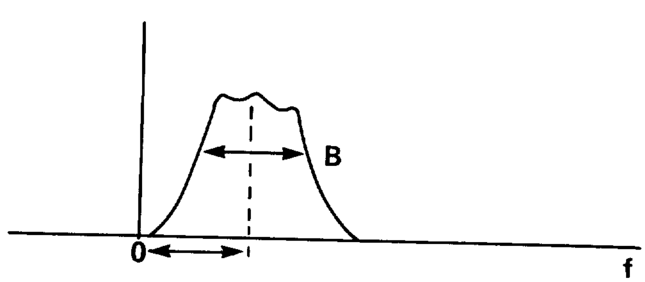

Fig. 6 Converted spectrum of mth channel

The chief disadvantage is that unless multiple bins are used (Fig. 1), the down-converted bandwidth may be too wide to capture the pulse on an oscilloscope or DRFM. As already noted, coherency is lost between the bins when the pulse is filtered into bins.

2.4. HOMOPULSE ENVELOPE DETECTION

Coherent detection can be used to measure the instantaneous envelope of a single pulse. This is accomplished by dividing the input pulse into two branches and then mixing (Fig. 7a). Let the pulses in the two branches be represented by the phasors (Fig. 7b).

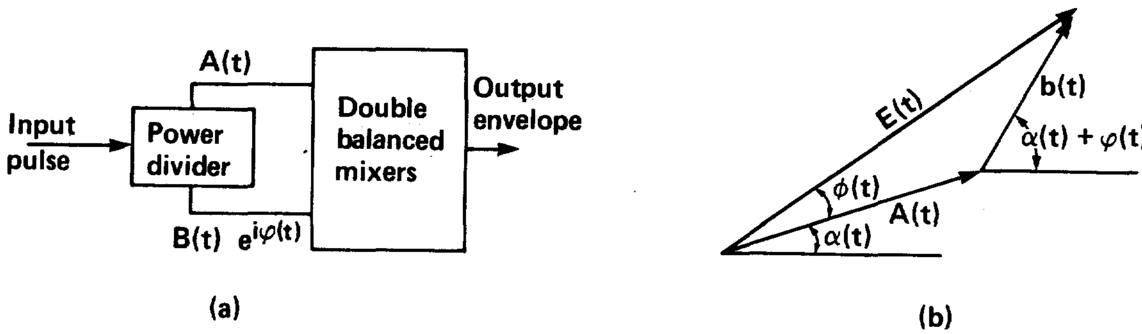

Fig. 7 (a) Homopulse envelope detector
(b) Phasor diagram of the two input branches to the mixer

$$e_A = A(t)\ e^{i\alpha(t)} \tag{6a}$$

$$e_B = B(t)\ e^{i[\alpha(t) + \phi(t)]} \tag{6b}$$

where A and B are the instantaneous amplitudes. The instantaneous frequency is $\omega(t) = \frac{d\alpha(t)}{dt}$.

For generality, $\phi(t)$ is included in the B-branch to permit a phase difference between the two branches, as may happen in the divider or mixer, or through unequal delay in the two branches.

Since the mixer is basically an envelope detector, the amplitude of the resulting phasor is

$$e_t^2 = A^2 + B^2 + 2AB \cos \phi(t) \tag{7}$$

by the law of cosines. If the sum of A and B is in the square law region, (7) represents the output of either one of the mixers. If the phase between the two balanced mixers is π, then the sign of either A or B is changed in one of the mixers, and the combined outputs of the two mixers are subtracted by the balancing process, yielding

$$e_t^2 = k(A^2 + B^2) + 2AB \cos \phi(t) \tag{8}$$

where the factor k is introduced to show the degree of imperfect balance. Typically a k-factor of about -17 to -25dB can be achieved. Thus, practically, the output is

$$e_t^2 \approx 2AB \cos \phi(t) \tag{9}$$

for square law mixers. Now, if the input power is divided equally, then A=B. However, note that the true <u>power</u> envelope (i.e., $A^2(t)$) is only achieved if the phases of two branches are perfectly balanced so that $\phi(t) = 0$. Clearly, non-zero $\phi(t)$ will reduce the output.

Certainly, quadrature phasing between the inputs to the two mixers should be avoided since $\phi = \pi/2$, making (9) zero. The case where there is a path length difference $\Delta\ell$ between the two branches also gives a nonzero $\phi(t)$ and can cause some distortion of the pulse shape. Then,

$$\phi(t) = 2\pi\Delta\ell f(t)/v_p \tag{10}$$

where f(t) is the instantaneous frequency. This frequency is likely to change over the pulse duration, causing the amplitude to vary along the pulse in accordance with (9).

Instead of square law mixing, high level or linear mixing is suggested. This case is more difficult to analyze for arbitrary $\phi(t)$. But when $\phi = 0$ then (7) is a perfect square of

$$e_t = A + B \tag{11}$$

and if $A = B$, $e_t = 2A(t)$ which is the desired voltage envelope of the pulse. The use of high level mixing [$-25 < (A+B) < +15$dBm] avoids square law operation at the low end and reverse self bias or damage to the mixer at the high end. This range can vary depending on the manufacturer and type of mixer.

3. FREQUENCY DIVISION TECHNIQUES

The recent development of microwave frequency dividers ("Halvers") opens the possibility of transferring very wide microwave bands in a coherent manner to the frequency region where high speed digital sampling is possible (Fig. 8). Microstrip dividing circuits are commercially available over octave bands up to 18 GHz, primarily for electronic countermeasure applications [3-5]. "Halvers," which operate on the principle of parametric subharmonic resonance, are coherent over the entire band. Consequently, the bandwidth is divided by exactly the same amount as the RF, i.e., 2^N where N is the number of stages. This is a very important advantage in contrast to heterodyne down-conversion where the bandwidth of the down-converted signal is the same as that of the original signal before mixing. In principle it should be possible to record the instantaneous frequency over a full active bandwidth. The operation is the inverse of the frequency multiplication process.

3.1. SIMPLE FREQUENCY DIVISION

Frequency dividers are rather simple in design (Fig. 9). Working designs in this range have recently been built and tested by the Canadian Department of National Defense [3], and are now commercially available [5].

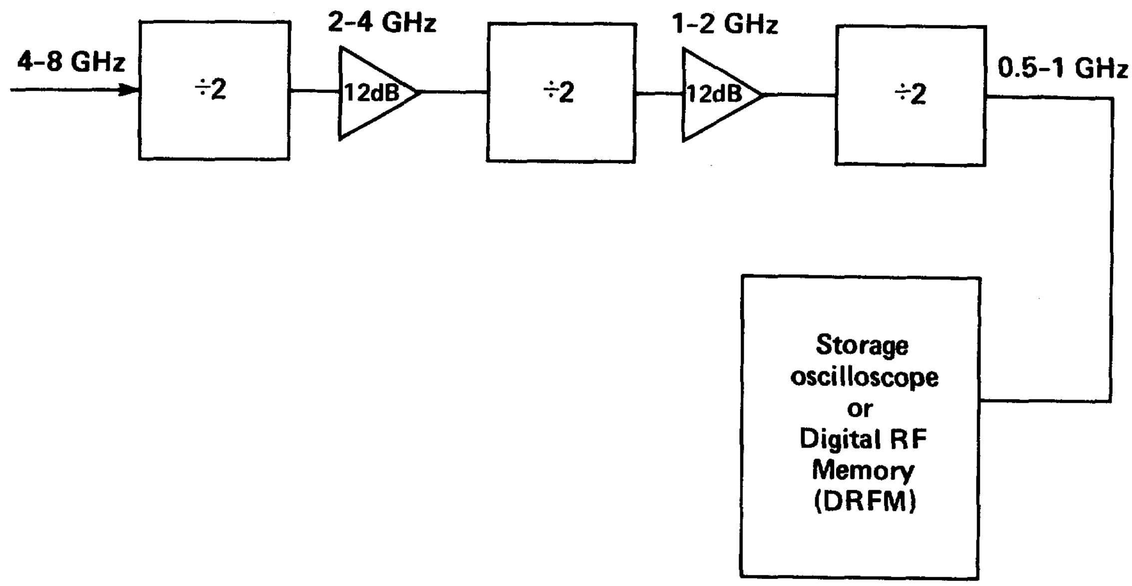

Fig. 8 Down-conversion and spectrum compression using frequency division

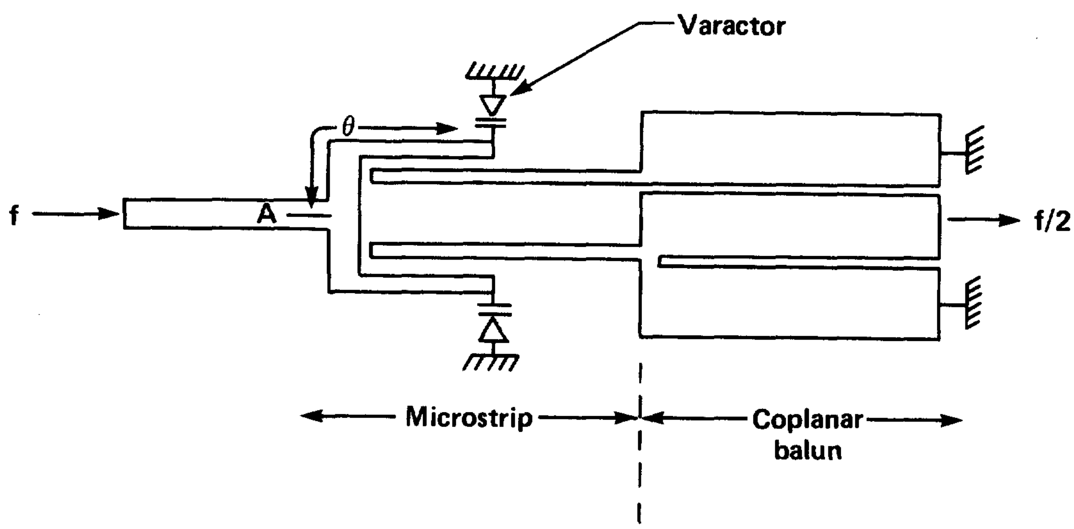

Fig. 9 Basic circuit layout of a frequency divider using microstrip and coplanar waveguide geometries

DC bias circuits for the diodes are not shown

Two important disadvantages of a "Halver" are that it tends to behave as limiter, and its high conversion loss. The limiting property causes an input 10dB dynamic range to be compressed into 2-3dB at the output, and also causes the device to respond to the strongest instantaneous signal. It therefore cannot be used to measure the instantaneous amplitude of a microwave burst, and its ability to measure the instantaneous frequency of exceedingly complex waveforms will require further research. To date, its primary use has been to compress such relatively well behaved signals as those corresponding to chirp or Doppler radars. The suppression phenomenon also causes spurious outputs when two frequencies occur simultaneously.

The conversion loss is typically 15dB, so 12-15dB of amplification is required between any two dividers. Halvers require the RF power to exceed a threshold before they "turn on." Consequently, the leading and trailing edges of the pulse will be lost, and the output appears as a "sharpened" pulse. Typically, an input of between 14 to 18dBm is required to turn the device on, although external biasing can be used to reduce the turn on threshold by 8-10dB.

Halvers typically turn on over a 10dB dynamic range. Overdriving causes some direct feed-through of the input frequency, but this can be suppressed using a low pass filter.

In spite of their limitations, halvers offer some unique potential for single pulse diagnostics, chiefly for reducing the frequency and bandwidth simultaneously. This deserves further attention. The following section shows one way in which the advantages of the halver can be retained while also overcoming its limiting properties.

3.2. FREQUENCY DIVISION/COHERENT DOWN CONVERSION

As noted in the preceding section, halvers tend to behave as limiters, so the output amplitude is not faithfully reproduced. The undesirable feature can be mitigated if the ouput of the halver is used as a coherent local oscillator and mixed with the original signal to achieve frequency division <u>through</u> down conversion, as shown in Fig. 10b. In this scheme, the input signal $Be^{i\alpha(t)}$ to be measured in both amplitude and frequency is split. The major portion goes through a limiter and the halver to generate a LO signal of $Ae^{i\frac{\alpha(t)}{2}}$ where

$$\omega_{1/2}(t) = \frac{1}{2}\frac{d\alpha(t)}{dt} \qquad\qquad 0 < t < \tau'$$

is the instantaneous frequency output by the halver, and τ' is the length of the halved pulse. Since the halver needs a threshold level to sustain its division function, the limiter is placed in front of the halver. Nevertheless, some pulse sharpening will occur so τ' is expected to be slightly less than the length of the input pulse, τ. The limiter together with the halver produces a constant pulse envelope of amplitude A. An amplifier is shown in Fig. 10a to compensate for the halver's conversion loss and to set the power level at the input to the mixer. Typically, A ~ 10dBm for high level mixing.

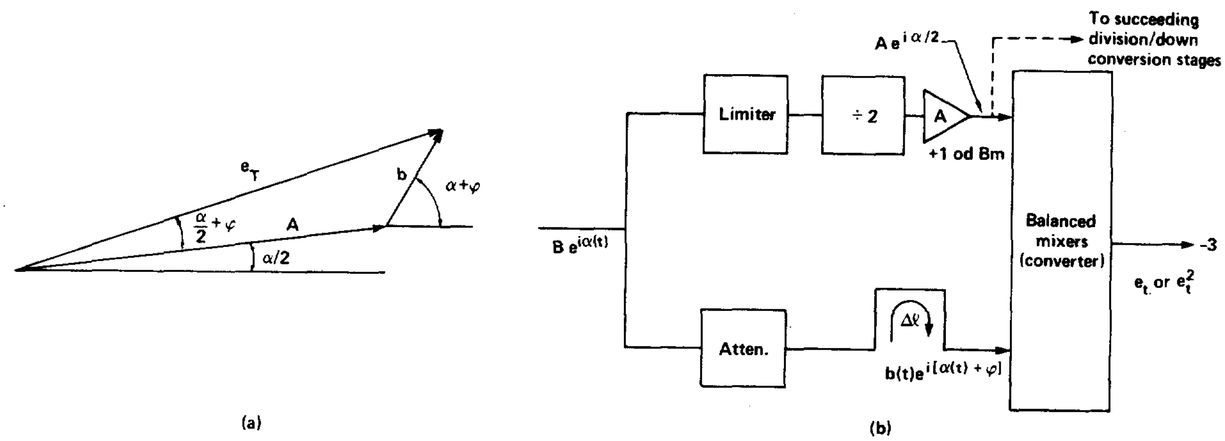

Fig. 10 First division/down conversion stage
(a) Phasor diagram of input to mixers
(b) Block diagram

The original input signal in the lower branch, b, is attenuated such that $b << A$, typically by 30dB. Some small delay, $\phi = 2\pi f\Delta\ell/v_p$, is added so that the true pulse arrives at the mixer at the same instant as the LO pulse, A.

Since the two signal inputs to the mixer are coherent to each other, a phasor diagram can be drawn during the time that both pulses exist (Fig. 10a). As before, the square of the total amplitude at the input to either one of the balanced mixers is

$$e_t^2 = A^2 + b^2 + 2Ab \cos\left[\frac{\alpha}{2} + \phi\right] \tag{12}$$

Mixers are envelope detectors, so the output of either mixer is proportional to (12), modified by the conversion loss. The balancing function suppresses the dc terms by the factor k (typically -17 to -25dB). The total output is then

$$e_T^2 = k\,(A^2 + b^2) + 2Ab \cos\left[\frac{\alpha}{2} + \phi\right] \tag{13}$$

where the conversion loss factor is omitted for clarity. If $b << A$ and the mixers operate in the square law region, the output is

$$e_T^2 = kA^2 + 2Ab \cos\left[\frac{\alpha}{2} + \phi\right] \tag{14}$$

This is a pulse of instantaneous frequency

$$\begin{aligned} \omega'_{1/2}(t) &= \frac{d}{dt}\left[\frac{\alpha(t)}{2} + \phi(t)\right] \\ &= \omega_{1/2}(t) + \frac{2\pi\Delta\ell}{\nu_p} f(t)\,\frac{df(t)}{dt} \end{aligned} \tag{15}$$

The second term in (15) represents a second order effect due to the small delay, $\Delta\ell$. The pulse in (14) is superimposed on a constant level kA^2. A coupling capacitor will allow the RF to pass and block the constant term.

For high level mixing where $A \sim 10$dBm, the mixers operate as a linear envelope detector. If $b << A$, then, with small error, the negligible second term in (12) can be multiplied by $\cos^2\left[\frac{\alpha}{2} + \phi\right]$ to create the perfect square of

$$e_t \cong A + b_{(t)} \cos\left[\frac{\alpha(t)}{2} + \phi\right] \tag{16}$$

which is the divided version of the input in which the desired amplitude information has been conserved. For balanced linear mixers, the constant term is suppressed by the factor k. Further suppression results from the use of a coupling capacitor which passes the RF.

The output RF pulse corresponding to (14) or (16), depending on whether the mixers are square law or linear, occurs for $0 < t < \tau'$. Since the true pulse extends beyond τ', the down conversion process will be lost for a short piece of the tail of the true pulse, i.e., for $\tau' < t < \tau$. For typical high power microwave sources, the energy in this lost piece is probably insignificant.

Multiple division/down conversion can be achieved by cascading N units like the single unit shown in Fig. 10b. This would accomplish division by 2^N. Figure 11 shows the second such unit. It is assumed that the amplitude of A from the first stage is sufficient to drive the second stage halver. Additional amplification is required after the second halver to compensate for its conversion loss and to provide enough signal to drive the second down converter and succeeding stages.

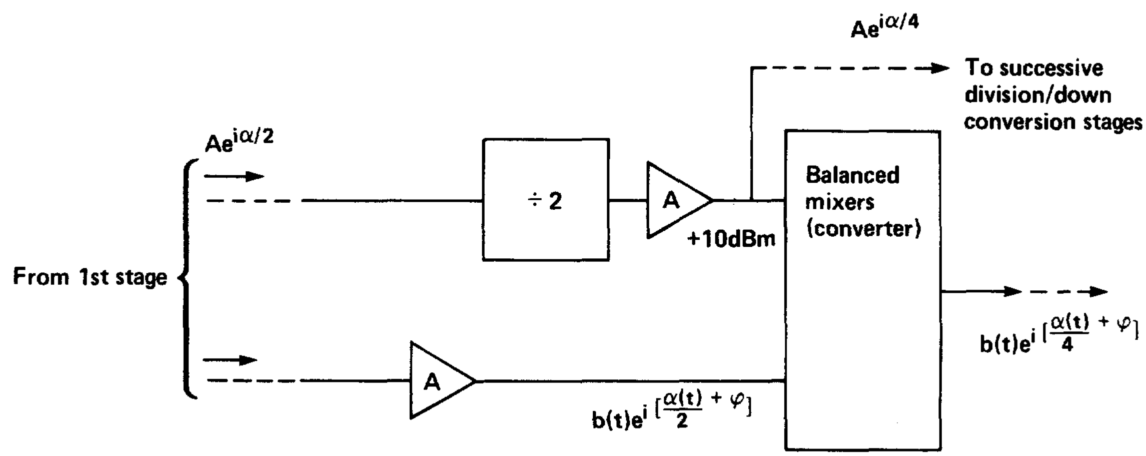

Fig. 11 Second division/down conversion stage

The output of the first stage converter may be amplified in the b-branch to compensate for the conversion loss.

After an appropriate number of stages, the resulting output can be captured and digitized and its spectrum analyzed.

The approach described above is offered for possible future research and development in single pulse high power diagnostics.

4. DIGITAL RF MEMORY (DRFM)

Primarily intended for applications in Electronic Countermeasures, high speed digital RF memories offer some promise of being able to capture microwave pulses up to 1 GHz [5]. Through the use of the heterodyne, homodyne or frequency division/techniques discussed in the preceding sections, DRFM's can and freeze the down converted pulse for several microseconds and recall for digital processing (e.g., further frequency division) and analysis.

The Telemus [5] sampler has a dynamic range of 40 dB and can produce in excess of 5 Gigabits/sec, sampled using apertures of 150 to 400 psec. If necessary, a custom sampler could be built which is capable of producing a 4 bit sample every 200 psec [6].

Besides use in capturing down-converted microwave pulses, DRFM's would also be useful for capturing single shot video pulses, e.g., simulated EMP and Enhanced HEMP threats.

REFERENCES

[1] Pershing, D., W. M. Bollen and R. Jackson, [1983], "Diagnostic Techniques for Short, Non-Repetitive High-Power Microwave Pulses," Mission Research Corp., P.O. Box 279, Springfield, VA 22150-0279. Technical Report DNA-TR-84-14.

[2] King, R. J., [1978], Microwave Homodyne Systems, Peregrinus Press, London.

[3] Cornish, W. D., (1982), "Microwave frequency dividers: devices and applications," IEE Proc. 129F(3), June, pp. 208-212.

[4] Harrison, R. G., (1977), "A broad-band frequency divider using microwave varactors," IEEE Trans. MTT-25(12), Dec., pp. 1055-1059.

[5] Telemus Electronic Systems, Inc., 310 Moodle Dr., Nepean, Ont. K2H 8G3, (613) 726-1102.

[6] Tucker, T. V., (1985), Pres. Telemus Electronic Systems (private communication).